(54) Title: BLIND CARRIER OFFSET DETECTION FOR QUADRATURE MODULATED DIGITAL COMMUNICATION SYSTEMS

(57) Abstract: A system and method for estimating the frequency offset experienced between carrier and local oscillator frequencies in communication systems using quadrature modulation. The invention exploits the geometry of the quadrature modulation constellation and estimates actual offset within a predefined carrier offset value without requiring data estimation.

BLIND CARRIER OFFSET DETECTION FOR QUADRATURE MODULATED DIGITAL COMMUNICATION SYSTEMS

BACKGROUND OF THE INVENTION

Field of the Invention

The present invention relates generally to digital communication systems using quadrature modulation techniques. More specifically, the invention relates to a system and method for blind detection of carrier frequency offsets in such systems.

Description of the Prior Art

A digital communication system typically transmits information or data using a continuous frequency carrier with modulation techniques that vary its amplitude, frequency or phase. After modulation, the signal is transmitted over a communication medium. The communication media may be guided or unguided, comprising copper, optical fiber or air and is commonly referred to as the communication channel.

The information to be transmitted is input in the form of a bit stream which is mapped onto a predetermined constellation that defines the modulation scheme. The mapping of each bit as symbols is referred to as modulation.

Each symbol transmitted in a symbol duration represents a unique waveform. The symbol rate or simply the rate of the system is the rate at which symbols are transmitted over the communication channel. A prior art digital communication system is shown in **FIG. 1**. While the communication system shown in **FIG. 1** shows a single communication link, those skilled in this art recognize that a plurality of multiple access protocols exist. Protocols such as frequency division multiple access (FDMA), time division multiple access (TDMA), carrier sense multiple

access (CSMA), code division multiple access (CDMA) and many others allow access to the same communication channel for more than one user. These techniques can be mixed together creating hybrid varieties of multiple access schemes such as time division duplex (TDD). The type of access protocol chosen is independent of the modulation type.

One family of modulation techniques is known as quadrature modulation and is based on two distinct waveforms that are orthogonal to each other. If two waveforms are transmitted simultaneously and do not interfere with each other, they are orthogonal. Two waveforms generally used for quadrature modulation are sine and cosine waveforms at the same frequency. The waveforms are defined as

$$s_1(t) = A\cos(2\pi f_c t) \qquad \text{Equation 1}$$

and

$$s_2(t) = A\sin(2\pi f_c t) \qquad \text{Equation 2}$$

where f_c is the carrier frequency of the modulated signal and A is the amplitude applied to both signals. The value of A is irrelevant to the operation of the system and is omitted in the discussion that follows. Each symbol in the modulation alphabet are linear combinations generated from the two basic waveforms and are of the form $a_1 \cos(2\pi f_c t) + a_2 \sin(2\pi f_c t)$ where a_1 and a_2 are real numbers. The symbols can be represented as complex numbers, $a_1 + ja_2$, where j is defined as $j = \sqrt{-1}$.

The waveforms of Equations 1 and 2 are the most common since all passband transmission systems, whether analog or digital, modulate the two waveforms with the original baseband data signal. Quadrature modulation schemes comprise various pulse amplitude modulation (PAM) schemes (where only one of the two basic waveforms is used), quadrature amplitude modulation (QAM) schemes, phase shift keying (PSK) modulation schemes, and others.

A prior art quadrature modulator is shown in **FIG. 2**. The modulator maps the input data as a pair of numbers $\{a_1, a_2\}$ which belong to a set defined by the modulation alphabet. a_1 represents the magnitude (scaling) of the first waveform and a_2 represents the magnitude (scaling) of the second waveform. Each magnitude is modulated (*i.e.* multiplied) by the orthogonal waveforms. Each individual modulator accepts two signal inputs and forms an output signal at the carrier frequency.

A prior art quadrature demodulator is shown in **FIG. 3**. The demodulator generates sine and cosine waves at a carrier frequency f_c for demodulation. Ignoring channel effects, the received signal can be represented as

$$r(t) = a_1(t)\cos(2\pi f_c t + \phi_0) + a_2(t)\sin(2\pi f_c t + \phi_0) \qquad \text{Equation 3}$$

where $a_1(t)$ represents the plurality of amplitudes modulated on waveform $s_1(t)$ as defined by Equation 1 and $a_2(t)$ represents the plurality of amplitudes modulated on waveform $s_2(t)$ as defined by Equation 2. ϕ_0 is an arbitrary phase offset which occurs during transmission.

The cosine and sine demodulator signal components are defined as:

$$
\begin{aligned}
r_c(t) &= r(t) * \cos(2\pi f_{LO} t) \\
&= \frac{1}{2} a_1 \cos((f_c - f_{LO})t + \phi_0) + \frac{1}{2} a_2 \sin((f_c - f_{LO})t + \phi_0) \qquad \text{Equation 4} \\
&+ \frac{1}{2} a_1 \cos((f_c + f_{LO})t + \phi_0) + \frac{1}{2} a_2 \sin((f_c - f_{LO})t + \phi_0)
\end{aligned}
$$

and

$$r_s(t) = r(t) * \sin(2\pi f_{LO}t)$$
$$= \frac{1}{2}a_2\cos((f_c - f_{LO})t + \phi_0) - \frac{1}{2}a_1\sin((f_c - f_{LO})t + \phi_0) \quad \text{Equation 5}$$
$$- \frac{1}{2}a_2\cos((f_c + f_{LO})t + \phi_0) + \frac{1}{2}a_1\sin((f_c - f_{LO})t + \phi_0)$$

The carrier frequency components, $f_c + f_{LO}$, are suppressed by the lowpass filters. The signals after filtering are:

$$y_c(t) = \frac{1}{2}a_1\cos((f_c - f_{LO})t + \phi_0) + \frac{1}{2}a_2\sin((f_c - f_{LO})t + \phi_0) \qquad \text{Equation 6}$$

and

$$y_s(t) = \frac{1}{2}a_2\cos((f_c - f_{LO})t + \phi_0) - \frac{1}{2}a_1\sin((f_c - f_{LO})t + \phi_0) \qquad \text{Equation 7}$$

If the local oscillator frequency in Equations 6 and 7 is equal to the carrier frequency, f_{LO} $= f_c$, and the phase offset is equal to zero, $\phi_o = 0$, the right hand sides of Equations 6 and 7 become ½ $a_1(t)$ and ½ $a_2(t)$ respectively. Therefore, to effect precise demodulation, the local oscillator must have the same frequency and phase as that of the carrier waveform. However, signal

perturbations occurring during transmission as well as frequency alignment errors between the local oscillators of the transmitter and receiver manifest a difference between the carrier and local oscillator frequencies which is known as carrier offset. A phase difference between the carrier and local oscillator frequency is created as well. However, if the difference in frequencies is corrected, the difference in phase is simple to remedy. Phase correction is beyond the scope of the present disclosure.

Carrier frequency offset is defined as:

$$\Delta f = f_c - f_{LO}.$$ Equation 8

To synchronize either parameter, the frequency and phase offsets need to be estimated. In prior art receivers, frequency offset estimation is performed after a significant amount of data processing. Without correcting offset first, the quality of downstream signal processing suffers.

What is needed is a system and method of detecting and estimating carrier frequency offset before any data signal processing is performed.

SUMMARY OF THE INVENTION

The present invention provides a system and method for estimating the frequency offset experienced between carrier and local oscillator frequencies in communication systems using quadrature modulation. The invention exploits the geometry of the quadrature modulation constellation and estimates actual offset within a predefined carrier offset value without requiring data estimation.

Accordingly, it is an object of the invention to provide a less-complex system and method for blindly estimating carrier frequency offset.

It is another object of the present invention to blindly estimate carrier offset in a communication system using quadrature modulation regardless of the access protocol.

Other objects and advantages of the system and method will become apparent to those

skilled in the art after reading a detailed description of the preferred embodiment.

BRIEF DESCRIPTION OF THE FIGURES

FIG. 1 is a simplified system diagram of a prior art digital communication system.

FIG. 2 is a system diagram of the prior art quadrature transmitter shown in FIG. 1.

FIG. 3 is a system diagram of the prior art quadrature receiver shown in FIG. 1.

FIG. 4 is a system diagram of the blind carrier offset estimator of the present invention.

FIG. 5 is a detailed system diagram of a blind digital carrier offset estimator of the present

invention.

DETAILED DESCRIPTION OF THE INVENTION

The embodiments will be described with reference to the drawing figures where like

numerals represent like elements throughout.

Shown in FIG. 4 is an analog or digital blind carrier detector 33 of the present invention.

A quadrature modulated signal $r(t)$ is received from a communication channel (not shown) and is

input 19 to a receiver 17. One skilled in this art recognizes that additional conversion means may

exist before the detector input 19 to convert the energy used in the transmission media to

compatible signals and is beyond the scope of this disclosure. The received signal $r(t)$ is coupled

to a cosine mixer 21_c and a sine mixer 21_s. Each mixer 21_c, 21_s has a first input 25_c, 25_s for

coupling with the received signal $r(t)$ and a second input 27_c, 27_s for coupling with the output of

a local oscillator LO. The local oscillator LO is programmed to generate cosine and sine waves

at the carrier frequency f_c (Equations 4 and 5) of the received signal $r(t)$.

The carrier-frequency demodulated outputs $r_c(t)$, $r_s(t)$ from each mixer 21_c, 21_s are input to respective lowpass filters 29_c, 29_s which suppress high-frequency noise components impressed upon the received signal $r(t)$ during transmission through the transmission media and mixer sum frequencies, $f_c + f_{LO}$, (Equations 6 and 7). As in prior art demodulators, the response characteristics of the lowpass filters 29_c, 29_s may be a bandwidth as narrow as Δf_{MAX} - the maximum allowable carrier offset. The output $y_c(t)$, $y_s(t)$ from each lowpass filter 29_c, 29_s is coupled to inputs 31_c, 31_s of a carrier offset estimator 33.

The carrier offset estimator 33 produces an estimate of the carrier offset 35 before data signal processing commences using a complex power processor 37 in conjunction with a complex Fourier transform processor 39. The filtered, carrier frequency demodulated cosine and sine components of the quadrature signal $y_c(t)$ and $y_s(t)$ are coupled to the complex power processor 37 which performs an intermediate power calculation of each quadrature component in the form of x^y where the powers y comprise integer multiples of four; i.e. $y = 4, 8, 12, 16 \ldots$. In the preferred embodiment, the power y is 4.

The complex power processor 37 may be implemented to raise the input complex signal to a power which is any positive integer multiple of four. Carrier offset detection systems which use a complex power processor with a power of two or its positive integer multiples are known in the art. However, these prior art systems do not work in quadrature-modulated digital communication systems. To properly detect a carrier offset in a quadrature-modulated digital communication system demodulator, a complex power of four or its integer multiples are necessary.

The complex power processor 37 combines the lowpass filter outputs $y_c(t)$ and $y_s(t)$ into a single complex value signal $y(t)$ defined as:

$$y(t) = y_c(t) + jy_s(t) \qquad\qquad \text{Equation 9}$$

where j is defined as $j = \sqrt{-1}$. The complex power processor **37** generates two power output signals

$$q_c(t) = Re\{(y(t))^4\} \qquad\qquad \text{Equation 10}$$

and

$$q_s(t) = Im\{(y(t))^4\} \qquad\qquad \text{Equation 11}$$

where $Re\{x\}$ denotes the real part of a complex number x, and $Im\{x\}$ denotes the imaginary part of the complex number x. The complex power processor **37** removes the modulation component from each received symbol leaving the carrier frequency. The real $q_c(t)$ and imaginary $q_s(t)$ signal components are output and coupled to the complex Fourier transform processor **39**.

The complex Fourier transform processor **39** treats the real $q_c(t)$ and imaginary $q_s(t)$ signal components as a single complex input signal $q(t) = q_c(t) + jq_s(t)$. The processor observes $q(t)$ for a finite period of time T_W and computes a complex Fourier transform of the observed signal $q(t)$ over this period of time.

The Fourier processor **39** performs a Fourier transform of the power processed signals from the observed period T_W and outputs a frequency at which the amplitude of the transform was measured to be maximal Δf_{MAX} during that time period T_W. The output **35** represents an accurate estimate of Δf and is signed since the transform input signal is complex. The sign identifies whether the local oscillator **LO** frequency is less than or greater than the carrier frequency.

A detailed, low-complexity digital implementation of the present invention **53** is shown in **FIG. 5**. Lowpass filter **29**$_c$, **29**$_s$ output signals $y_c(t)$ and $y_s(t)$ are sampled at a sampling rate f_s to produce discrete-time signals $y_c[n]$ and $y_s[n]$. To ensure that all possible carrier frequency offsets up to Δf_{MAX} are detected, $2\Delta f_{MAX} < f_s$ must be satisfied. The passband of the low pass

filters 29_c, 29_s must be wider than Δf_{MAX} to avoid suppressing the signal which contains the carrier offset information.

The sampled signals $y_c[n]$ and $y_s[n]$ are input 51_c, 51_s to a complex power processor **57** and combined as a single complex signal, $y[n]$, where $y[n] = y_c[n] + j y_s[n]$. The power processor **57** produces a complex output defined by $q[n] = (y[n])^4$. The output $q[n]$ is coupled to a buffer **59** for accumulating N outputs from the complex power processor **57**.

The accumulated block of complex numbers N is coupled to a digital Fourier transform (DFT) processor **61** which performs a transform from the time domain to the frequency domain for the N complex numbers. The DFT processor **61** outputs N complex numbers corresponding with the input N. Each number is associated with a particular frequency ranging from $-f_s/2$ to $(+f_s/2 - f_s/N)$. Each frequency is f_s/N away from a neighboring frequency. The frequency domain values output by the DFT **61** are assembled and compared with one another. The value having the largest magnitude represents the best estimate of the carrier frequency offset Δf.

The embodiment described in **FIG. 5** is capable of estimating all carrier frequency offsets smaller than $f_s/2$. This follows from the restriction $2\Delta f_{MAX} < f_s$ imposed above. The carrier offset Δf is resolved to within a frequency uncertainty of $\pm f_s/2N$ since the frequencies at the output of the DFT **61** are quantized to a grid with a spacing of f_s/N. Since the frequencies are f_s/N away from each other, the invention **53** renders precision within $\pm\frac{1}{2}$ of the selected value. Therefore, the number of samples N accumulated for the Fourier processor **61** to transform determines the resolution of the carrier offset estimate Δf. An efficient implementation of the DFT **61** used in the present invention **53** can be achieved using the fast Fourier transforms (FFT) family of algorithms.

The present invention **33, 53** may be physically realized as digital hardware or as software. The lowpass filters shown in **FIG. 5** may be realized in digital hardware or software

operating at a sampling rate faster than f_s. In some communication systems, for example those employing CDMA protocols, the lowpass filters and downsamplers f_s may be replaced with accumulators and integrate-and-dump processes.

While the present invention has been described in terms of the preferred embodiments, other variations which are within the scope of the invention as outlined in the claims below will be apparent to those skilled in the art.

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CLAIMS

What is claimed is:

1. A detector (33) for use in a receiver (17) that estimates a difference (Δf) between a received quadrature demodulated communication signal ($r(t)$) carrier frequency (f_c) and a local oscillator (LO) frequency used for demodulation, the detector (33) characterized by:

a complex power processor (37) for receiving the quadrature demodulated communication signal ($r(t)$) cosine ($y_c(t)$) and sine ($y_s(t)$) components producing real ($Re\{(y(t))^p\}$) and imaginary ($Im\{(y(t))^p\}$) power signal components, said power signal components coupled to a complex Fourier transform (39);

said complex Fourier transform (39) performing a complex Fourier transform on said power signal components for a predefined period of time outputting a frequency transform containing a plurality of frequencies observed during said predefined period of time; and

said plurality of frequencies having a maximum frequency corresponding to the difference (Δf) between the carrier (f_c) and local oscillator (LO) frequencies.

2. The detector (33) according to claim 1, wherein said complex Fourier transform (39) maximum frequency is further characterized with a sign ($\pm$) indicating whether the difference frequency (Δf) is greater than or less than the local oscillator (LO) frequency.

3. The detector (33) according to claim 2, wherein said complex power processor (37) is further characterized by obtaining a power (p) that is an integer multiple of four.

4. A method (33) for estimating a difference (Δf) between a received (17) quadrature demodulated communication signal ($r(t)$) carrier frequency (f_c) and a local oscillator (LO) frequency used for demodulation, the method (33) characterized by the steps of:

a) performing a complex power calculation on the received quadrature demodulated communication signal ($r(t)$) cosine ($y_c(t)$) and sine ($y_s(t)$) components producing real ($Re\{(y(t))^\gamma\}$) and imaginary ($Im\{(y(t))^\gamma\}$) power signal components;

b) transforming said real ($Re\{(y(t))^\gamma\}$) and imaginary ($Im\{(y(t))^\gamma\}$) power signal components into Fourier frequency domain values representing a plurality of frequencies;

c) selecting a maximum frequency from said plurality of frequencies; and

d) outputting said selected frequency as the difference frequency (Δf).

5. The method (33) according to claim 4 wherein step d) is further characterized by a step of assigning a sign ($\pm$) to said maximum frequency indicating whether the offset frequency (Δf) is greater than or less than the local oscillator (LO) frequency.

6. The method (33) according to claim 5 wherein step a) is further characterized by a step of obtaining powers ($^\gamma$) that are integer multiples of four.

7. A detector (53) for use in a receiver (17) that estimates a difference (Δf) between a received quadrature demodulated communication signal ($r(t)$) carrier frequency (f_c) and a local oscillator (LO) frequency used for demodulation, the detector (53) characterized by:

a sampler (f_s) sampling the received quadrature demodulated communication signal ($r(t)$) signal components ($y_c(t)$), ($y_s(t)$) into discrete time ($y_c[n]$)), ($y_s[n]$) signal components, said discrete time ($y_c[n]$), ($y_s[n]$) signal components coupled to a complex power processor (57);

said complex power processor (57) performs a complex power calculation on said discrete time ($y_c[n]$), ($y_s[n]$) signal components producing real ($Re\{(y[n])^y\}$) and imaginary ($Im\{(y[n])^y\}$) power signal components which are output ($q[n]$) to a buffer (59);

said buffer (59) accumulates a plurality (N) of complex power processor (57) outputs ($q[n]$) for a predefined period of time as a block of data (T_w), said block of data (T_w) is output to a complex Fourier processor (61);

said complex Fourier transform (61) performs a complex Fourier transform on said block of data (T_w) outputting a frequency transform containing a plurality of frequencies ($-f_s/2$ to ($+f_s/2 - f_s/N$)) observed during said predefined period of time, said output ($-f_s/2$ to ($+f_s/2 - f_s/N$)) is coupled to a selector; and

said selector (63) selects from said plurality of frequencies a maximum frequency which corresponds to the difference (Δf) between the carrier (f_c) and local oscillator (LO) frequencies.

8. The detector (53) according to claim 7, wherein said sampler (f_s) is further characterized by sampling the received quadrature demodulated continuous time ($y_c(t)$), ($y_s(t)$) signal components at a predefined ($2\Delta f_{MAX} < f_s$) frequency.

9. The detector (53) according to claim 8, wherein said complex Fourier transform (39) maximum frequency is further characterized with a sign ($\pm$) indicating whether the difference frequency (Δf) is greater than or less than the local oscillator (LO) frequency.

10. The detector (53) according to claim 9, wherein said complex power processor (57) is further characterized by obtaining a power (y) that is an integer multiple of four.

11. The detector (53) according to claim 10, wherein said complex Fourier transform (39) is further characterized as a fast Fourier transform.

12. A method (53) for estimating a difference (Δf) between a received (17) quadrature demodulated communication signal ($r(t)$) carrier frequency (f_c) and a local oscillator (LO) frequency used for demodulation, the method (53) characterized by the steps of:

a) sampling (f_s) received ($r(t)$) quadrature demodulated communication signal ($r(t)$) signal components ($y_c(t)$), ($y_s(t)$) into discrete time ($y_c[n]$)), ($y_s[n]$) signal components;

b) performing a complex power calculation (57) on said discrete time ($y_c[n]$), ($y_s[n]$) signal components;

c) buffering (59) said complex power calculation output (($y[n]$)') for a plurality (N) of complex power processor (57) outputs for a predefined period of time as a block of data (T_w);

d) transforming (61) said block of data (T_w) into Fourier frequency domain values representing a plurality of frequencies ($-f_s/2$ to ($+f_s/2 - f_s/N$));

e) selecting (63) a maximum frequency from said plurality of frequencies ($-f_s/2$ to ($+f_s/2 - f_s/N$)); and

f) outputting said selected frequency as the difference frequency (Δf).

13. The detector (53) according to claim 12, wherein step a) is further characterized by a step of sampling the received quadrature demodulated continuous time ($y_c(t)$), ($y_s(t)$) signal components at a predefined ($2\Delta f_{MAX} < f_s$) frequency.

14. The detector (53) according to claim 13, wherein step d) is further characterized by a step of assigning a sign (±) to said maximum frequency indicating whether the offset frequency (Δf) is greater than or less than the local oscillator (LO) frequency.

15. The detector (53) according to claim 14, wherein step b) is further characterized by a step of obtaining a power (P) that is an integer multiple of four.

16. The detector (53) according to claim 15, wherein step d) is further characterized as a fast Fourier transform.
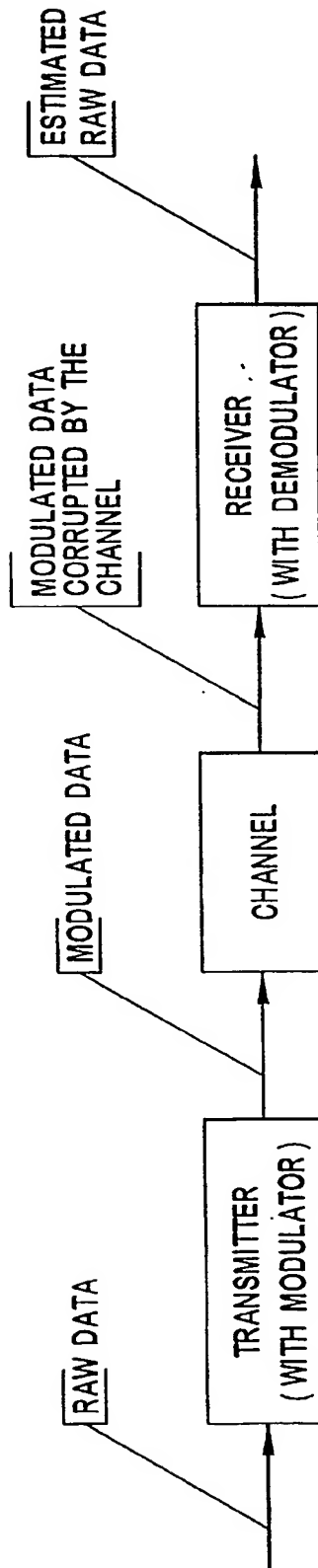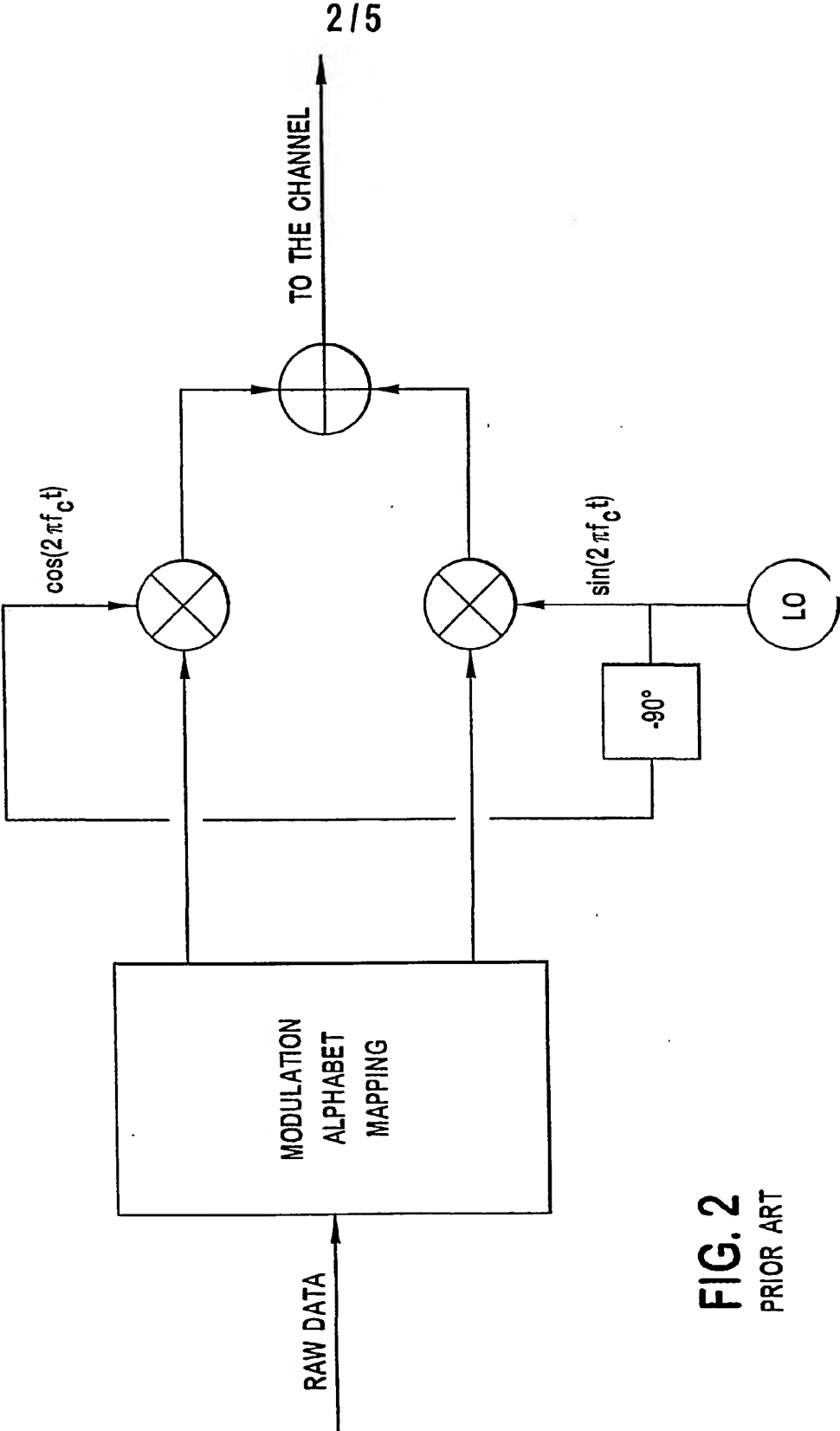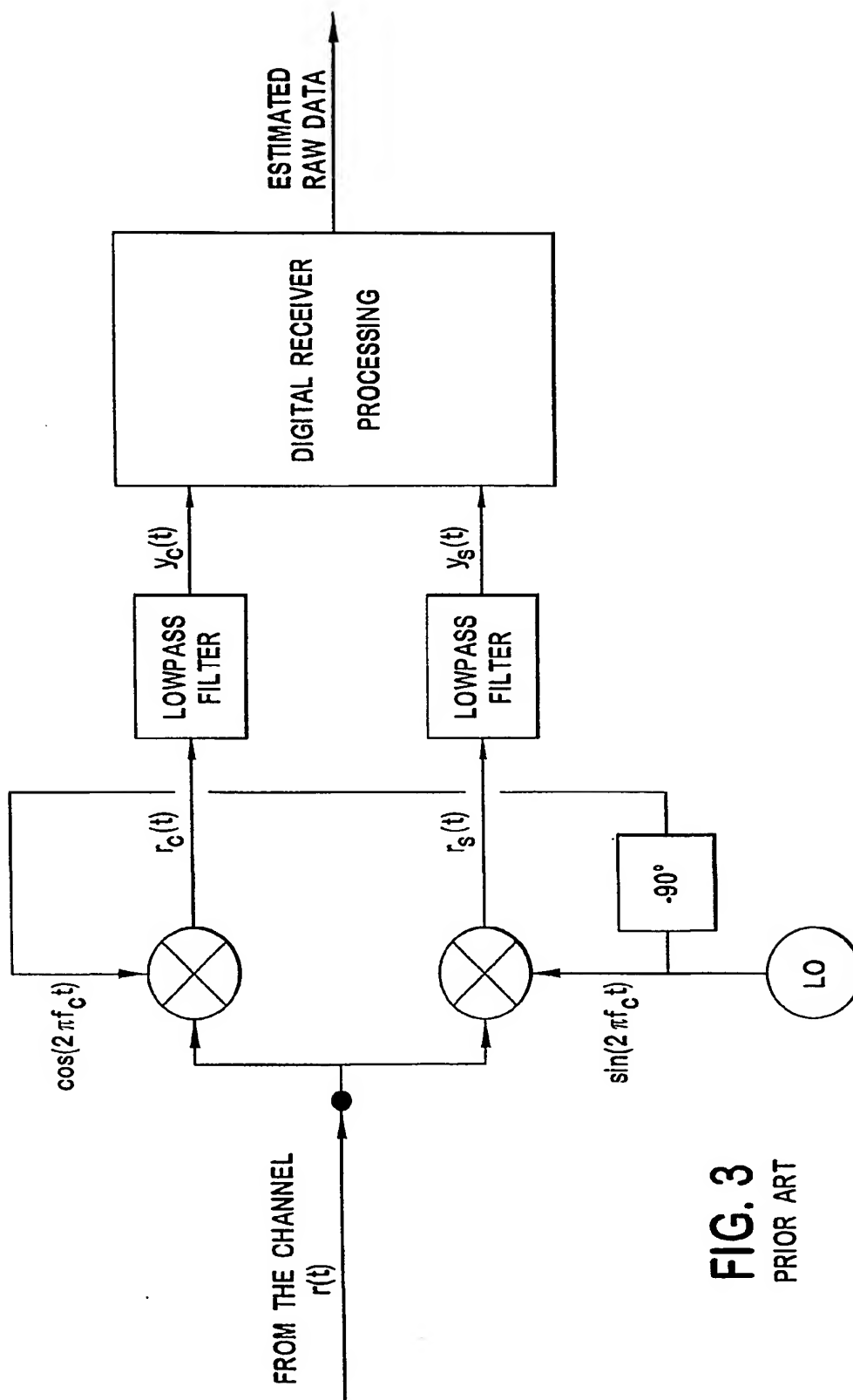
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FIG. 1
PRIOR ART

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FIG. 2
PRIOR ART

FIG. 3
PRIOR ART

FIGURES 4 AND 5

NOT FURNISHED UPON FILING

NO PRESENTADO(A) EN EL MOMENTO DE LA PRESENTACIÓN

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A. CLASSIFICATION OF SUBJECT MATTER
IPC 7   H04L27/233   H04L27/38

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7   H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	COWLEY W G ET AL: "ESTIMATION OF FREQUENCY OFFSET IN MOBILE SATELLITE MODEMS" INTERNATIONAL MOBILE SATELLITE CONFERENCE, 16 - 18 June 1993, pages 417-422, XP000978070 PASADENA, CALIFORNIA, USA page 417, column 1, line 1 -page 421, column 2, line 7; figures 1,3-5 --- -/--	1-16

| X | Further documents are listed in the continuation of box C. | | Patent family members are listed in annex. |

* Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

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Date of the actual completion of the international search	Date of mailing of the international search report
2 January 2001	16/01/2001

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C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT		
Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
Y	FRANKS L E: "CARRIER AND BIT SYNCHRONIZATION IN DATA COMMUNICATION A TUTORIAL REVIEW" IEEE TRANSACTIONS ON COMMUNICATIONS,US,IEEE INC. NEW YORK, vol. COM-28, no. 8, 1 August 1980 (1980-08-01), pages 1107-1121, XP000565125 ISSN: 0090-6778 page 1108, column 2, line 23 -page 1108, column 2, line 40; figure 1 page 1110, column 2, line 29 -page 1111, column 1, line 30	1-16
Y	AHMED W K M ET AL: "A METHOD FOR COARSE FREQUENCY ACQUISITION FOR NYQUIST FILTERED MPSK" IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY,US,IEEE INC. NEW YORK, vol. 45, no. 4, 1 November 1996 (1996-11-01), pages 720-731, XP000681449 ISSN: 0018-9545 page 720, column 1, line 1 -page 722, column 1, line 27; figure 1	1-16
A	TAUB H ET SCHILLING D L: "PRINCIPLES OF COMMUNICATION SYSTEMS" 1986 , MCGRAW-HILL , NEW YORK XP002155414 page 264, line 13 -page 265, line 37; figure 6.5.5 page 275, line 1 -page 276, line 3; figure 6.7.3	1-16